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COMPREHENSIVE FLUX ESTIMATOR IMPLEMENTATION PROCEDURES FOR ADVANCED CONTROL OF INVERTER-FED INDUCTION MACHINES

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Abstract— **This paper presents the implementation of a flux estimator for induction machines. Both stator and rotor flux are estimated using the back electromotive force method. A procedure for the implementation, calibration, and testing of this estimator on a digital signal processor is given. The procedure is intended for applications in which the estimator is essential but not a primary system development target. The work presented here can be extended to other estimators, especially for advanced inverter-fed motor control applications.**

Key Words: Induction Motor Drives, Flux Estimator, DSP Implementation Procedure

I. INTRODUCTION

The trend in motor drives for the last three decades has been to implement vector methods such as field-oriented control (FOC) [1] and direct torque control (DTC) [2]. Advanced control techniques such as feedback linearization [3] are also of interest. Open-loop or V/Hz control is still found in several applications such as small pumps and fans, but drives that require higher performance standards tend to use advanced control methods where energy-saving and optimization methods can also be applied [4, 5]. Several commercial motor drives incorporate vector-based controls to satisfy the growing market demand for higher performance and more efficient drives. Most available control techniques for induction machines require information about the flux. For example, FOC requires stator or rotor flux estimation in the synchronous frame, DTC requires stator flux estimation in the stationary frame, feedback linearization requires rotor flux estimation in the synchronous frame, etc. A typical induction motor drive is shown in Figure 1.

The work presented here looks at the estimator as a "gray box" within the Control Estimation block of Fig. 1. The details of the estimation process are not the purpose of the study, rather general estimator characteristics such as inputs, outputs, estimation error, calibration, and implementation are thoroughly discussed. The back electromotive force (EMF) estimator was chosen because it is basic and relatively easy to follow for analysis and implementation.

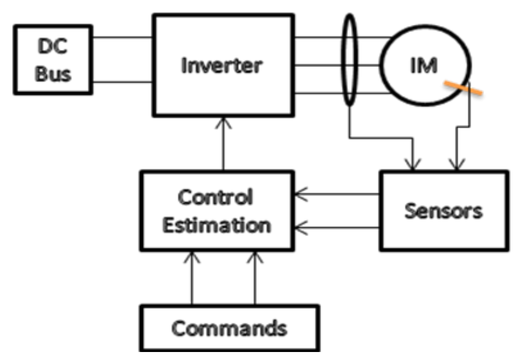

Figure 1: Typical induction motor drive

Figure 2 shows a flowchart summarizing the estimator implementation procedure. A similar procedure was used to implement V/Hz and FOC controllers, and this procedure can be extended to other motor control or power electronics applications with a similar digital control platform. The main steps of the procedure are the preparation of hardware and software, interfacing, implementation, calibration, and testing.

II. FLUX ESTIMATORS

Flux estimators can be categorized into three groups according to [6]: back EMF methods [6-13], model reference adaptive systems (MRAS) [14, 15], and observer-based approaches [16-23]. These estimators can have several variations under each category. The list of categories can be extended to include artificial neural networks [24]. Most of the available literature focuses on the ability of estimators to reduce errors in the magnitude and phase of the estimated flux relative to the real flux. Back EMF estimators are usually based on

$$\lambda^s_{qd,s} = \int \left(v^s_{qd,s} - R_s i^s_{qd,s} \right) dt \qquad (1)$$

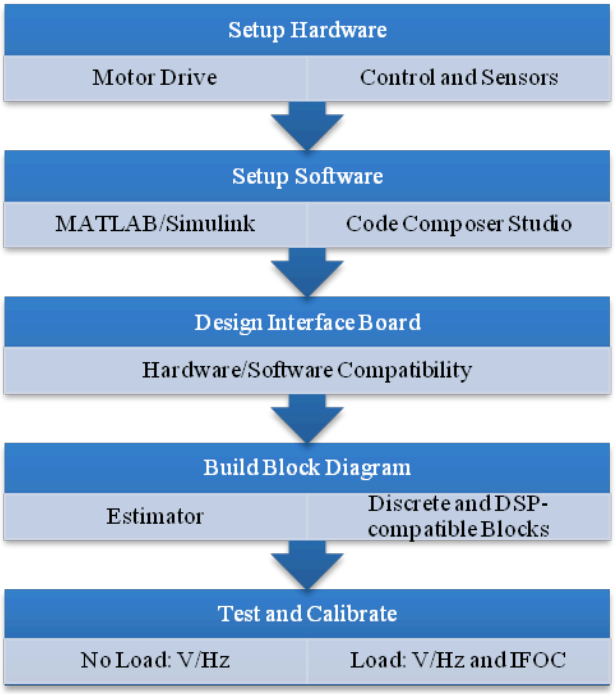

Figure 2: Implementation procedure of the flux estimator

In (1), $\lambda^s_{qd,s}$ is the stator flux linkage in the direct axis, $v^s_{qd,s}$ and $i^s_{qd,s}$ are the associated voltage and current, and R_s is the stator resistance. These variables are referenced to the stator in the stationary frame, in which voltages and currents can be measured directly. Eq. (1) might result in a dc offset owing to small errors in the voltage and current measurements [11]. The integrator would accumulate these errors and might saturate analog devices or cause an overflow in digital registers. One of the suggested solutions is to use a low-pass filter (LPF) instead of an integrator. An LPF will cause both magnitude and phase errors that need compensation [11].

Estimators based on Kalman filters linearize the machine model or restructure the model into a state-space form. Nonlinear observers are more complex and can be designed to produce better estimates [25]. Both sliding-mode observers [21, 23], and Luenberger observers [16, 17, 20] have been presented for flux estimation applications. Further background on observer-based flux estimators is available in [26].

While most of the available estimators target the stator flux, rotor flux estimates are also essential in indirect FOC (IFOC). Thus, several rotor-flux estimators have been proposed. A straightforward estimator based on (1) is

$$\lambda^s_{qd,r} = \frac{L_r}{L_m}\left(\lambda^s_{qd,s} - \sigma L_s i^s_{qd,s}\right) \quad (2)$$

where $\lambda^s_{qd,r}$ is the rotor flux linkage, L_m is the magnetizing inductance, L_r is the rotor inductance, and σ is a leakage factor defined as $\sigma = 1 - L_m^2/L_r L_s$. The estimate from (2) suffers from the same drawbacks as the back EMF estimator since it is a linear function of $\lambda^s_{qd,s}$. A popular estimator with corrective abilities is presented in [27]. This estimator is given by

$$G = \left(I - K\frac{L_m}{L_r}\right)\lambda^s_{qd,r} - K\sigma L_s i^s_{qd,s} \quad (3)$$

$$\lambda^s_{qd,r} = \left(I - K\frac{L_m}{L_r}\right)^{-1}\left(G + K\sigma L_s i^s_{qd,s}\right) \quad (4)$$

where G is an intermediate iteration variable, I is a 2x2 identity matrix, and K is a 2x2 diagonal estimator gain. The literature does not include a detailed procedure for implementing and calibrating a flux estimator. For example, a digital implementation was employed in [11], but the complete strategy is missing there and the work is difficult to reproduce.

The procedure presented here targets a back EMF estimator and is divided into four main sections. Section III includes details about the hardware and software requirements. Section IV explains the method of implementing and debugging the estimator. Calibration, testing, and results are presented in Section V.

III. Software and Hardware Requirements

The work presented here employs a conventional digital control platform. It is based on the eZdsp F2812™ board as a suitable platform for implementing motor controllers. This board is built around the TMS320F2812 digital signal processor (DSP). This platform is compatible with Simulink®, and includes six dual pulse width modulation (PWM) channels (12 channels total), 16 analog to digital converters (ADC), and a speed encoder input [28]. The processor is a 32-bit DSP with fixed-point arithmetic; thus, discrete and fixed-point math blocks from Simulink can be used to program it.

A. Software Requirements

Two primary software packages must be available on the host computer where the development and control take place: MATLAB®/Simulink, which support math and control development, and Code Composer Studio (CCS), which supports detailed code development for the DSP. Compatibility is essential. For example, MATLAB 7.0.4 must be used with CCS 2.21, MATLAB 2006a must be used with CCS 3.1, etc. Simulink provides a simple user interface where a designer can build the estimator using discrete-time blocks and special DSP-related blocks from the "C2000" library, such as the fixed-point math "C28x IQmath" library.

Simulink is able to compile a block diagram into C code and then call CCS to generate assembly code for the DSP. A project is generated in CCS to be loaded into the DSP. MATLAB can also be used to build a friendly guided user interface (GUI) for real-time communication with the DSP through its parallel port using real-time data exchange channels (RTDX). These channels are set in the block diagram to be assigned on the DSP. Figure 3 shows a summary of the setup. It is important to note that programmers with experience in C or assembly languages can write their own optimized code for the DSP with basic tools, but this is a time-consuming process.

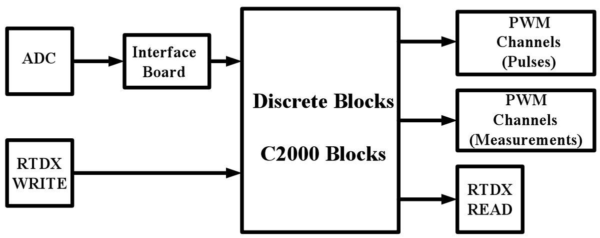

Figure 3: Hardware and software interconnections

B. Hardware Requirements

The only special hardware requirements for flux estimation are the host computer and the DSP board. The eZdsp used here is also able to operate offline from the computer, since it operates from an independent power supply. Current and voltage sensors are built into the power stage of the inverter in most advanced motor drives. The procedure presented here can be easily modified to other DSPs used as motor-control platforms. The sensors required by the estimator vary depending on the application and the estimation process. For example, DTC requires the stator flux estimate in the stationary frame; thus stator voltages and currents are necessary. For IFOC, voltage sensors might not be required, but a speed encoder is needed. Each advanced control technique has a minimum sensor requirement. The requirements are not likely to exceed 3 voltage sensors, 4 current sensors with one measuring the neutral current, and a speed encoder.

The interface board shown in Figure 3 may be needed, depending on the compatibility of the sensors with the DSP. This board takes the sensor outputs and conditions them for the DSP. For example, the eZdsp F2812 requires all inputs to the ADC to be between 0 and 3 V [28]. While simple voltage dividers with limited currents are straightforward, many current sensors have dc offsets and special input/output relations. Figure 4 shows a sample current-sensor characteristic and the expected output of the interface board fed into the eZdsp. The scaling circuitry usually consists of simple op-amp adders, gains, and filters. Additional digital filters on the DSP may be used to eliminate small offsets.

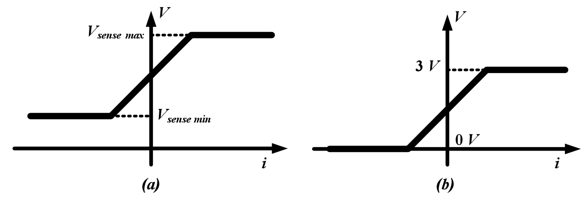

Figure 4: (a) Sensor output, (b) Scaled sensor output compatible with the DSP

IV. Implementation Procedure

After the sensors are scaled and conditioned for the ADC, the designer can read hardware information into Simulink. The outputs from Simulink are usually the PWM channels on the DSP — six dual channels, three of which are used to control the three-phase inverter, leaving three for measurements. The PWM outputs are discrete-valued, but the underlying modulation signal can be extracted with appropriate filtering. Other measurements can be made using the RTDX where the signals are sent from the DSP to the computer. RTDX can either read or write; thus, with an appropriate GUI, the designer can measure and send commands to the DSP in real time. The target eZdsp, and sample ADC, PWM, and RTDX blocks from Simulink are shown in Figure 5.

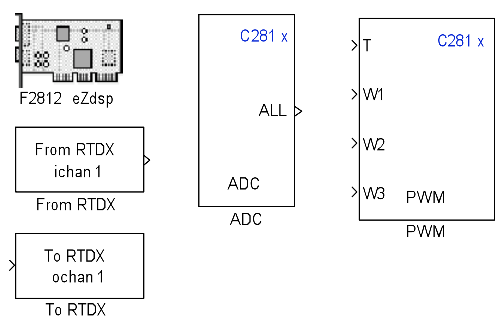

Figure 5: Sample DSP-related blocks in Simulink

When the hardware and MATLAB/Simulink and CCS are installed, building the estimator can start. The target hardware needs to be selected from the "C2000 Target Preferences" library. An important step is to set a discrete step size for the simulator. In general, smaller step sizes are better, provided the computational burden can be managed. In this system, the DSP clock frequency is 150MHz. In the flux estimator application, a sampling time of T_s=30 μs is feasible with this clock rate. The ADC has two modules, A and B, which can be accessed using the "ADC" block found in the "C28x DSP Chip Support" library. The sampling rate of the ADC can go up to 12.5 MHz, depending on the number of channels being read. The flux estimator here uses 7 channels; thus a low sampling rate should be used [29]. It is important to mention that the encoder input is separate from the ADC on this DSP.

The ADC output is an array that must be de-multiplexed to recover the individual signals. Its data type is and unsigned 16-bit integer. Only twelve bits are used by the ADC, however, where an analog level of 3 V corresponds to 4096, the full-scale 12-bit representation. This poses the need for appropriate software scaling. An example is shown in Figure 6. A current of 20 A translates to 5 V, before scaling to 3 V for DSP compatibility. The internal scale factor must be 20/4096 to translate the ADC output into current at 20 A full scale.

To achieve the best decimal accuracy, fixed-point math is used on the DSP. The “Target for TI C2000” Simulink library has the “C28x IQmath” module in which basic accurate math operations can be found [30]. The Q value of an IQ number is the bit before which the decimal point is placed. IQMATH blocks accept signed 32-bit fixed-point integers (sfix(32)). Figure 7 shows a sample of an sfix(32) number with Q=29. All data types must be converted to sfix(32) numbers to be implemented on the DSP. A detailed description of the IQMATH library is available in [30].

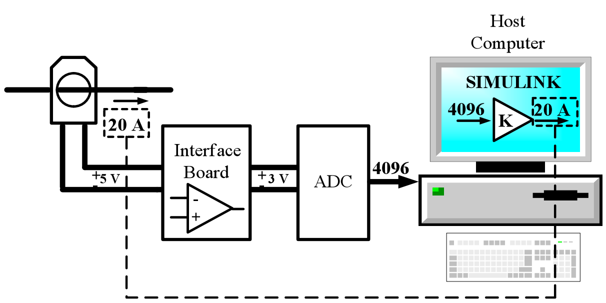

Figure 6: An example of a scaling procedure of a sensor output

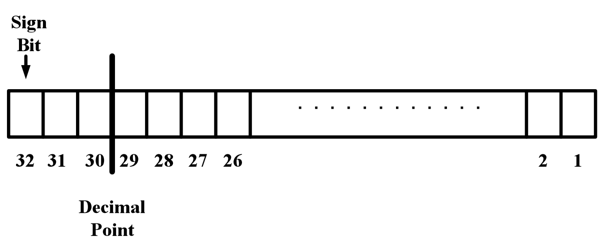

Figure 7: 32-bit signed fixed-point number

The implementation of (1) is straightforward except for the integrator. Several back EMF estimators replace it with a LPF because of the integrator dc offset. We used a discrete-time integrator sampled at T_s, and as expected, the dc offset existed and caused the integrator output to grow without bound. To solve this problem, the scheme shown in Figure 8 was implemented, where z is a one-unit delay. A LPF is added to extract the dc offset, which is first reduced by the high-pass filter (HPF).

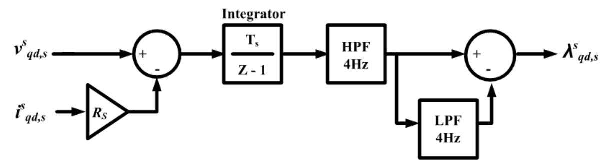

Figure 8: Modified flux estimator

The rotor flux is estimated using (2). All constants need to have the sfix(32) data type for accurate values. For example, the leakage term, σ, has a value close to 1, so Q=30 is used to represent it. On the other hand, the stator voltages have significantly higher values and can be represented with a smaller Q. The rotor flux in the synchronous frame is found by using Park’s transformation [1]. The *cosine* and *sine* terms need to have Q=30 for highest precision since their values are between -1 and 1. An example of an IQmath application is the rotor angle ρ found using the “Arctangent IQN” block, based on

$$\rho = tan^{-1}\left(\frac{\lambda^s_{q,r}}{\lambda^s_{d,r}}\right) \quad (5)$$

To further improve the estimation and achieve a higher accuracy, the flux is estimated in mV·s. This gives more room for decimal bits in the 32-bit number.

V. Testing and Results

The flux estimator described in (1) and (2) was implemented with a scheme similar to the one shown in Figure 8. All the blocks utilized in Simulink were as described in the previous sections, and the hardware setup is shown in Figure 9. The eZdsp board and the interface board are shown in Figure 10.

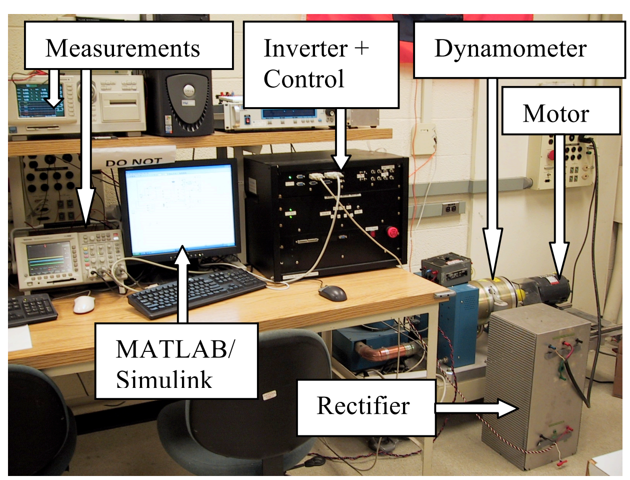

Figure 9: Hardware setup

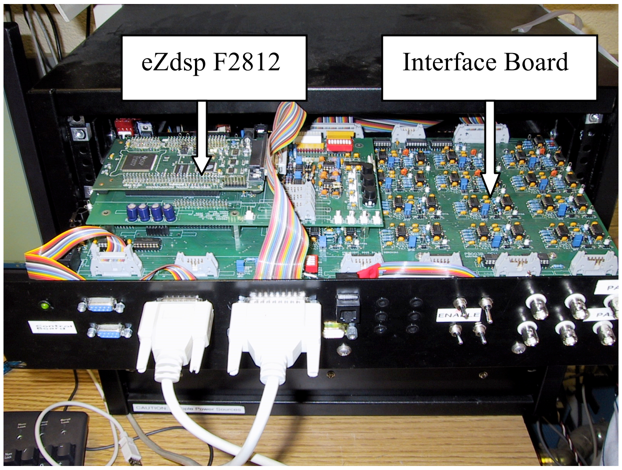

Figure 10:

The fluxes computed by the estimator were fed into RTDX channels to be plotted and saved in real time. The results need to be checked for accuracy against a reference flux.

The first procedure simulates the motor drive and estimator in Simulink, with the same step size as the sampling rate of the hardware. Significant errors existed between the hardware and simulation. The main issues are:

1) The use of an ideal simulation model.
2) Quantization errors.

The simulation model used was ideal in several ways. No saturation or core loss models were added, which affected the flux-current relationships. The inverter was simulated as an ideal amplifier, and the lines interconnecting the system were lossless. Also, noise effects, which might be significant, were ignored. Quantization errors can cause significant problems, especially with the integrator. An example is shown in Figure 11 where part of a sine wave is drawn. To reduce the quantization error, a higher ADC sampling rate can be used, without violating the sampling limit [29].

To account for the non-ideal real-life situation, another simulation approach was used. The rms terminal voltages and line currents of the motor were measured from experiments. The measured values in effect take into account all losses and nonlinearities. These results were then fed into a simulation implementing (1) and (2) at a high sampling rate, resembling continuous time relative to T_s. The results from this simulation give a clear idea about the quantization errors in the system. The sampling time was 25 µs for V/Hz and 30 µs for IFOC in simulations and experiments. The switching frequency for the V/Hz was 10 kHz; IFOC used hysteretic control.

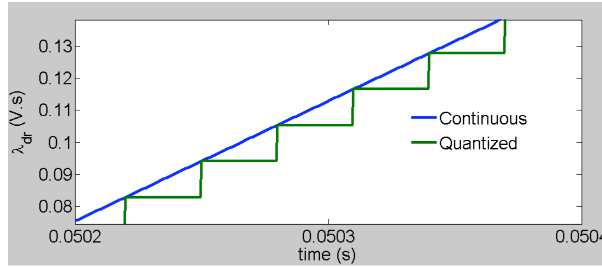

Figure 11: Quantized versus continuous flux

The main comparison between the simulations and experiments was made under V/Hz with no load. The other tests were made with a load under V/Hz and IFOC.

A. V/Hz with No Load

The stator current is low, and the effect of the stator voltage drop is negligible. This reflects a relatively ideal V/Hz control with low losses. This test has a predictable peak stator flux; thus experimental and simulation results can be compared to theory if needed. Table 1 summarizes the peaks of the stator and rotor fluxes in simulations (Sim.) and experiments (Exp.), for low and high speeds. Figures 12 and 13 show the simulation and hardware results. The errors in Table 1 are very low and reflect an acceptable estimation. The error could be a result of quantization errors and inaccuracies in motor parameters.

Table 1: Flux peaks from simulations and experiments under V/Hz with no load

	1226 rpm (42Hz)			600 rpm (20.7Hz)		
	Sim.	Exp.	Error	Sim.	Exp.	Error
$\lambda^s_{d,s}$	415	400	3.6%	325	350	7.7%
$\lambda^s_{d,r}$	410	390	4.9%	320	310	3.2%

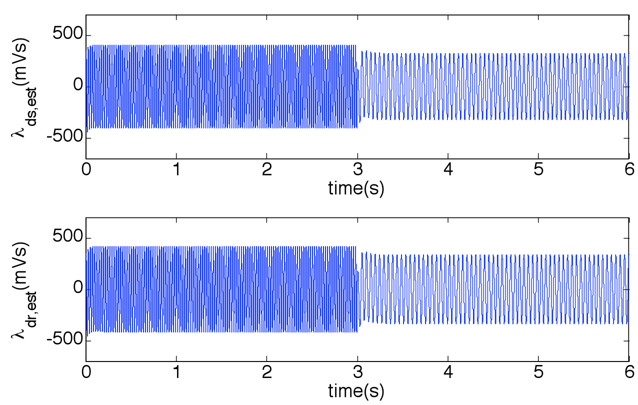

Figure 12: Simulation results for d-axis stator and rotor fluxes under V/Hz with no load

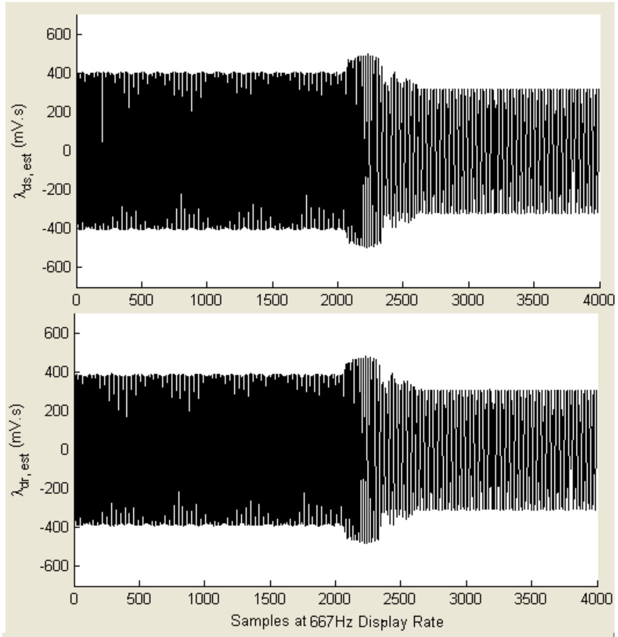

Figure 13: Hardware results for d-axis stator and rotor fluxes under V/Hz with no load

It is time-consuming to synchronize the hardware and simulations to consider the phase shift,. Thus, a simple approach is to predict the phase shift due to the first-order HPF at 4Hz with transfer function *H(jω)* where *ω* is the frequency in rad/s. For example, if the flux is at 60Hz, the phase shift, *φ*, can be predicted as

$$H(j\omega) = \frac{j\omega}{j\omega + 2\pi 4} \quad (6)$$

$$\varphi = tan^{-1}(2\pi 60) - tan^{-1}\left(\frac{2\pi 60}{2\pi 4}\right) = 89.58^{o} \quad (7)$$

The phase shift due to the LPF is negligible with about 1.54^{o}. Thus the effect of the HPF can be easily predicted and compensated for by about 90^{o}.

B. V/Hz with Load

To account for the effects of loads and higher losses that result under loading conditions on the flux estimator, the motor was loaded under different speeds. At a speed of 1226 rpm, the load torque was 3N·m, while at 600 rpm, the load torque was 0.7N·m. A similar procedure to the no-load case was followed, and the results are shown in Table 2 and Figures 14 and 15. Figure 16 shows the experimental torque and speed plots. The results show that the flux estimation error under load is higher but is still around 10%. The main issue is believed to be parameter error especially because the rotor time constant and the stator resistance vary under load where the motor temperature increases.

Table 2: Flux peaks from simulations and experiments under V/Hz with load

	1226 rpm (42Hz), 3N·m			600 rpm (20.7Hz), 0.7N·m		
	Sim.	Exp.	Error	Sim.	Exp.	Error
$\lambda^{s}_{d,s}$	355	315	11.3%	301	290	3.7%
$\lambda^{s}_{d,r}$	345	305	11.6%	295	265	10.2%

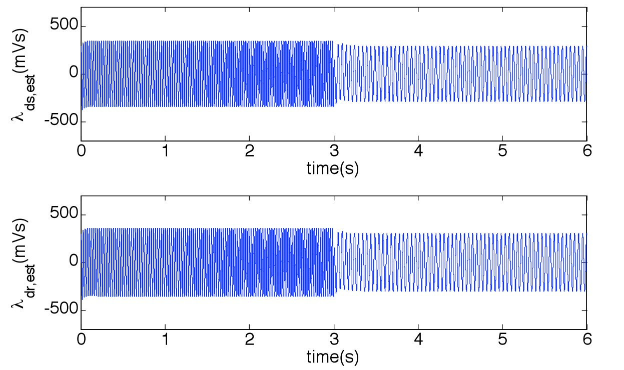

Figure 14: Simulation results for d-axis stator and rotor fluxes under V/Hz with load

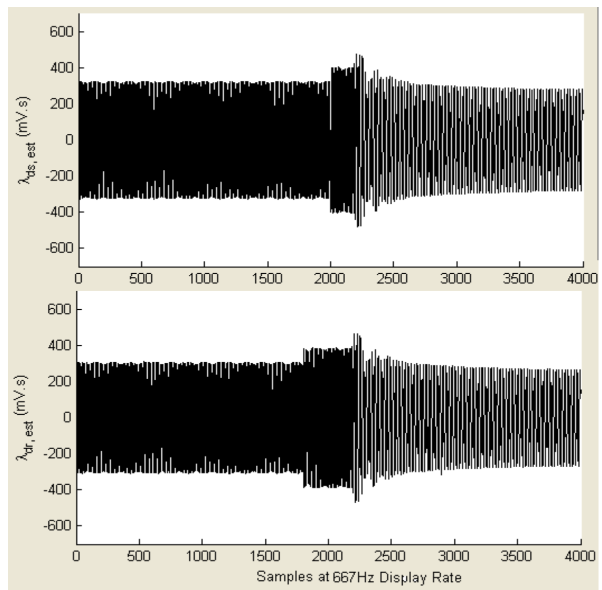

Figure 15: Hardware results for d-axis stator and rotor fluxes under V/Hz with load

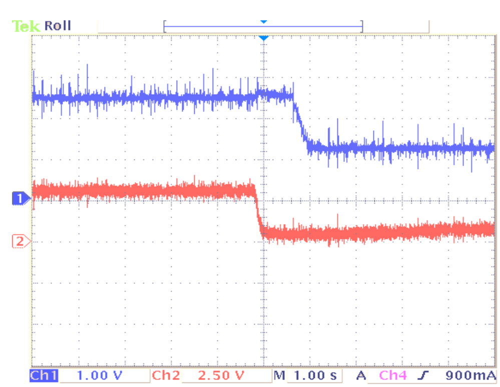

Figure 16: Speed (Ch1: 500rpm/div) and Torque (Ch2: 2N.m/div) under V/Hz

C. IFOC with Load

The last test with IFOC requires a flux estimate. The command flux is 500 mV·s. This test needs special calibration due to the noise and uncontrolled switching under hysteresis. The currents and voltages under IFOC have higher harmonic contents than fixed PWM in V/Hz control; therefore, quantization errors are more significant. Running the estimator in simulation and hardware without any calibration resulted in estimation errors of more than 30%. To compensate for quantization errors and noise, simple calibration was needed. A constant gain of 8/7 was used in the stator estimator to achieve better flux estimates. The value of the gain was determined by running the simulations and hardware, then comparing the peaks and scaling the stator flux. Table 3 shows the errors in the peaks from experiments and simulations, and Figures 17 and 18 show the actual and estimated fluxes. Figure 19 shows the experimental torque and speed plots.

The results in Table 3 have a low error rate even though the estimation is performed under hysteretic control. The calibration of the gain is straightforward and can be further improved. For example, a look-up table of gains can be used for different operating conditions.

Table 3: Flux peaks from simulations and experiments under IFOC with load

	1226 rpm (42Hz), 3N·m Load			600 rpm (20.7Hz), 0.7N·m Load		
	Sim.	Exp.	Error	Sim.	Exp.	Error
$\lambda^s_{d,s}$	394	400	1.5%	460	430	7%
$\lambda^s_{d,r}$	384	385	0.26%	448	415	8%

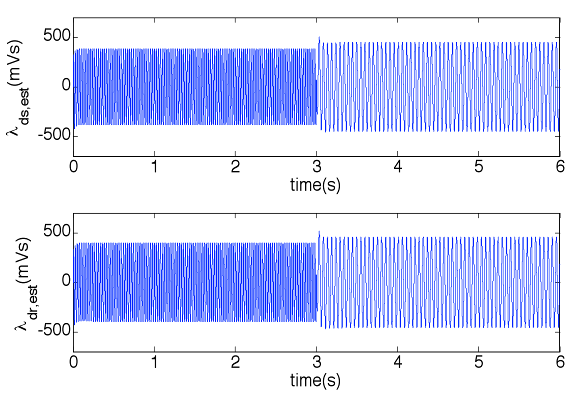

Figure 17: Simulation results for d-axis stator and rotor fluxes under IFOC with load

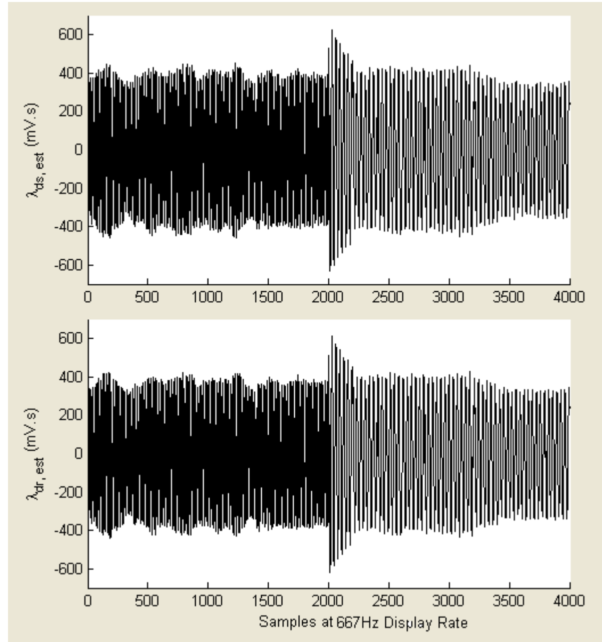

Figure 18: Hardware results for d-axis stator and rotor fluxes under IFOC with load

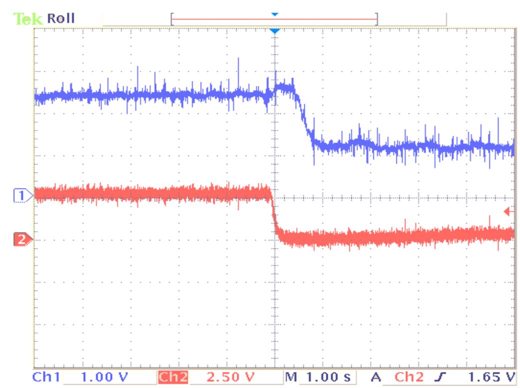

Figure 19: Speed (Ch1: 500rpm/div, upper trace) and Torque (Ch2: 2N.m/div, lower trace) under IFOC

VI. Conclusion

A comprehensive procedure for the implementation of a flux estimator was presented. The procedure is based on using a DSP as a control and estimation platform for inverter-fed induction machines. Hardware and software requirements, interfacing procedures, implementation, calibration, and testing were all presented. Detailed but simplified discussions of the fixed-point processing requirements for motor drives and other power electronics applications were also presented. This procedure can be used for several other applications where real-time DSP control and estimation are required.

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